A FIXED-GAIN, LOW-DISTORTION AF AMPLIFIER

by G. J. Estep

INTRODUCTION

The availability of small-signal, fixed-gain, monolithic amplifiers capable of delivering moderate output current has been limited by inflexible production techniques. In the past, various methods of obtaining useful output current capability have been tried, such as by using hybrids or monolithic operational amplifiers in conjunction with discrete output transistors. Although these methods have been successful, their use is generally restricted to those applications where cost is of secondary importance.

In general, the problems associated with the production of monolithic amplifiers designed for ac operation are more tractable than in the dc amplifier case. Since typical production variations are a fact of life and since a very stable gain is desired, the use of a feedback system is justified. The feedback concept provides excellent control over the production variations typically encountered, and allows a low-distortion, fixed-gain amplifier to be produced at very low cost (compared to presently available monolithic IC's).

The μA716 is designed for applications requiring a stable, fixed-gain device with voltage gains of 200, 100, 20, and 10. The design goals included an operating supply voltage range of 18 to 24 volts with a power output capability in excess of 100 mW when operated from a nominal 21-volt supply. In addition, external capacitive coupling to the load was desired to allow the use of smaller isolation transformers at the output of the system. Single supply operation was anticipated and bias currents were not allowed in the transformer windings since the cost of the transformer, in addition to its size, would be increased. The output impedance of the amplifier was to be adjustable to within 10 percent of some nominal value, typically 600 ohms. Operation over the entire audio spectrum with at least 40 decibels of feedback was also desired.

CIRCUIT DESCRIPTION

To satisfy the low-distortion requirement of the amplifier, an emitter-coupled pair was utilized for the input stage (Fig. 1). This configuration allows a reasonably high voltage gain with very low distortion for typical input voltages. The emitter-coupled pair also provides a convenient high-impedance, inverting input for the feedback network. In Fig. 1, the feedback network is a simple voltage divider connected to the base of Q_4. Since this node impedance is quite high (of the order of 400 kilohms), negligible loading of the feedback network is realized. Connection of a loop decoupling capacitor to the appropriate terminal of the amplifier allows any one of the four gain options to be utilized (Appendix II).

Since a true Darlington connection reduces the voltage gain of the input stage by a factor of two, resistors R_4 and R_5 are introduced between the base and emitter of Q_2 and Q_3, respectively. These resistors provide sufficient pull-down current to cause the source impedance (as seen by Q_2 and Q_3) to become negligible (Appendix I). Use of resistors rather than current sources is justified since common mode rejection is not a parameter of particular concern.

Voltage gain of the input stage is given approximately by the ratio of the effective load resistance and the sum of dynamic and contact resistances of devices Q_2 and Q_3. The effect of the source impedance is assumed negligible. Emitter follower, Q_5, decreases the loading of the collector resistance, R_6, so that the simplified gain expression is

$$A_V = \frac{R'_L}{2(r_e + R_c)} \qquad (1)$$

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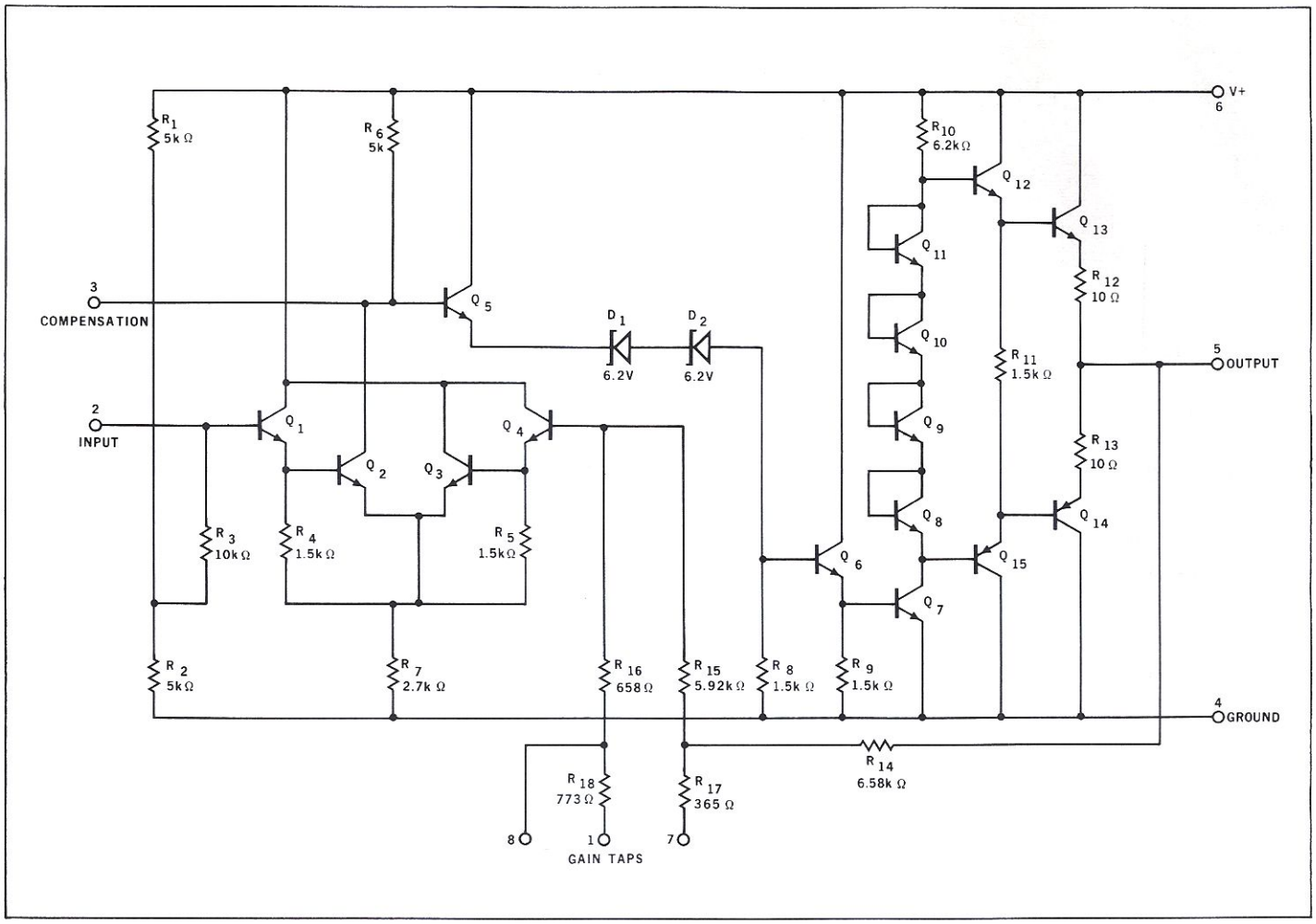


Fig. 1. Circuit diagram of integrated amplifier.

where

R'_L = load resistance of Q_2 including effect of level shifter

r_e = dynamic emitter resistance of Q_2 or Q_3

R_c = emitter contact resistance of Q_2 or Q_3.

This expression is true only for matched devices with equal collector currents in Q_2 and Q_3. Since these conditions are design conditions, some departure from symmetry is to be expected with production devices. (A complete expression for the voltage gain of the amplifier has been developed but is not necessary at this point. The results of the detailed evaluation of voltage gain will be summarized in a later section.)

The simultaneous achievement of low-distortion, low standby current, and large load current capability are well provided by the complementary, Class AB output stage. The diode-connected transistors offer excellent matching to the output devices so that the threshold is typically a few millivolts (open loop), and in addition, their dynamic resistance is considerably below that of an equivalent linear resistor, which means the voltage gain of the stage is determined essentially by R_{10}. The use of complementary devices insure that the load is driven by a low impedance source during positive and negative signal excursions. The essential distortion mechanism in the output stage is the degradation in current gain of the output devices for large currents. Insofar as the closed-loop amplifier is concerned, typical distortion for the highest available gain of 200 is on the order of 0.5 percent. Resistors R_{12} and R_{13} are provided to absorb any variation in bias voltage that may arise, and to prevent thermal runaway of the output devices for large internal dissipation. Standby current in Q_{13} and Q_{14} is typically 1 millampere and is less than 2 millamperes for worst-case mismatch in transistor and diode forward voltage.

Level translation from the input stage to the output stage is accomplished by the use of Zener diodes, D_1 and D_2, and emitter-followers, Q_5 and Q_6. The emitter-followers reduce the loading of the input stage and provide temperature stabliization of the translated voltage.

Voltage gain of the output stage and the level shifter is given by an expression similar to that of the input stage.

Addition of a suitable constant to account for the non-ideal shift of unity is all that is necessary, thus

$$A_v = \frac{kR'_{10}}{(r_e + R_c)} \qquad (2)$$

where

k = level shifter gain

R'_{10} = effective collector load on Q_7

$r_e + R_c$ = sum of emitter dynamic resistance and contact resistance.

OVERALL VOLTAGE GAIN

As mentioned in the previous section, the overall voltage gain of the general purpose amplifier is a function of several varying parameters. The more dominant variations are those connected with the emitter dynamic resistance, transistor current gain and base-emitter voltage, and integrated resistors. Unfortunately, most of the parameters of concern are not well defined in terms of temperature stability, so that empirical relationships must be determined from measured data. The one exception, of course, is the base-emitter voltage and the emitter-dynamic resistance which follow the well-known diode equation. Little can be said concerning the initial value and the temperature stability of the emitter contact resistance.

A detailed analysis of the amplifier voltage gain has been performed on the SDS-930 digital computer which included the variations in the above parameters. Empirically derived expressions for resistor variation and beta variation were used in the analysis. Emitter contact resistance was taken as a parameter in the plots of voltage gain versus temperature, as were the initial betas of the transistors. The general expression for the voltage gain is

$$A_v = (k)\,(gm_1)\,(gm_2)\,(R_6)\,(R_{10}), \qquad (3)$$

where

gm_1 = transconductance of input stage

gm_2 = transconductance of output stage

R_6 = effective load on input stage

R_{10} = effective load on output stage

k = level shifter gain constant.

The value of the above terms varies with temperature and supply voltage. In addition, the amplifier load resistor determines, to some extent, the gain of the amplifier, since beta of the output transistors is a function of output current.

The results of six sets of calculations are shown in Figs. 2 through 7 for emitter contact resistances of 0, 3.5, and 10 ohms and PNP room temperature betas of 30 and 10. For these calculations, room temperature beta of the NPN

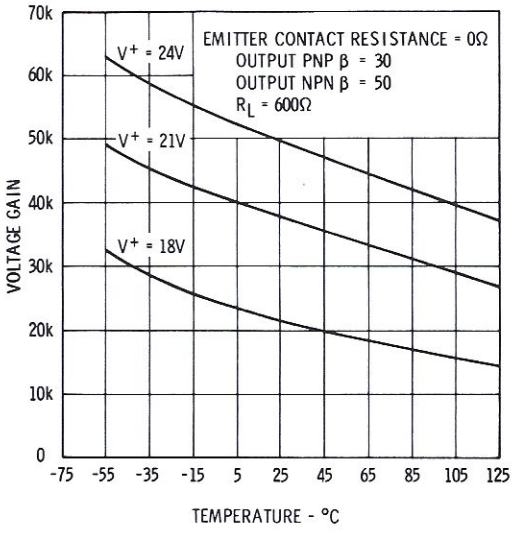


Fig. 2. Voltage gain as a function of ambient temperature.

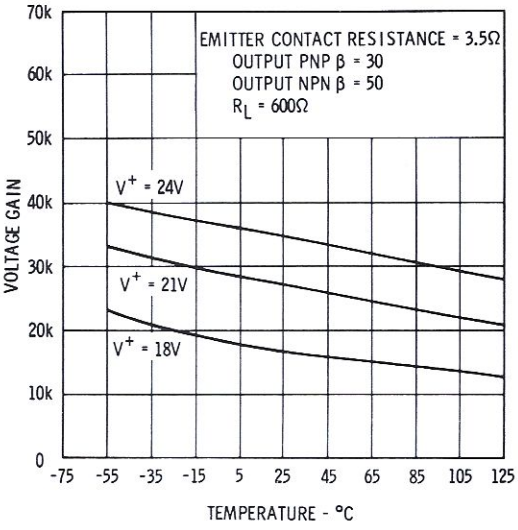


Fig. 3. Voltage gain as a function of ambient temperature.

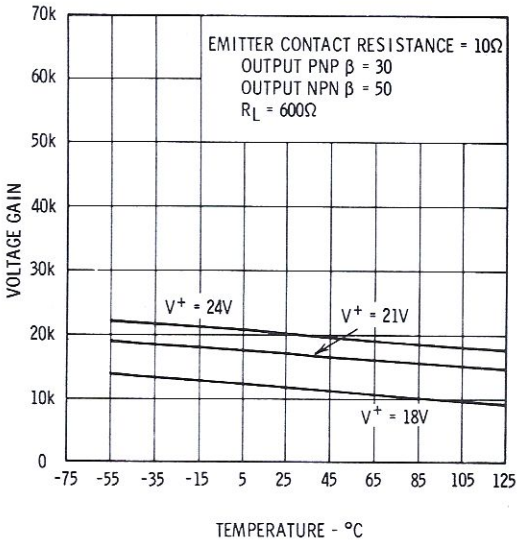


Fig. 4. Voltage gain as a function of ambient temperature.

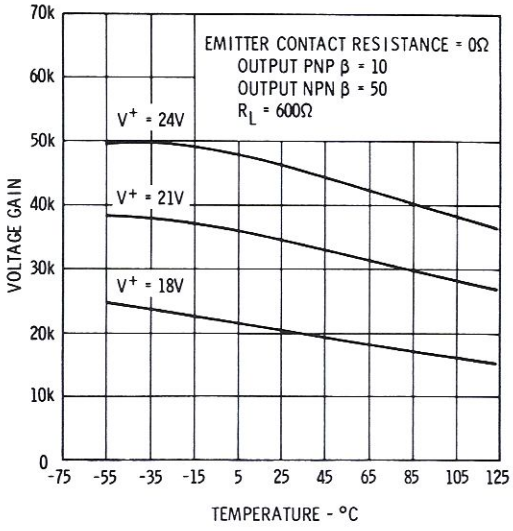


Fig. 5. Voltage gain as a function of ambient temperature.

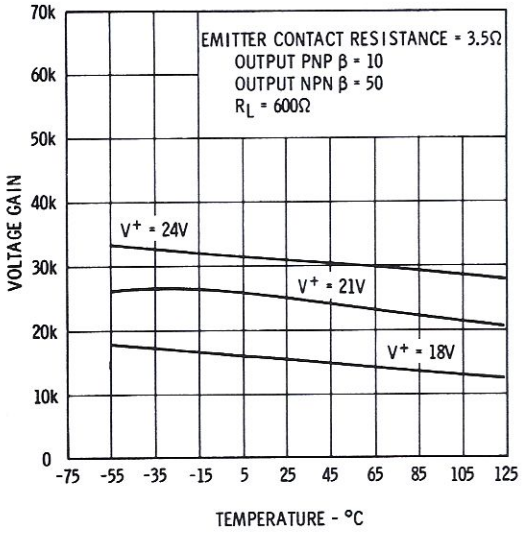


Fig. 6. Voltage gain as a function of ambient temperature.

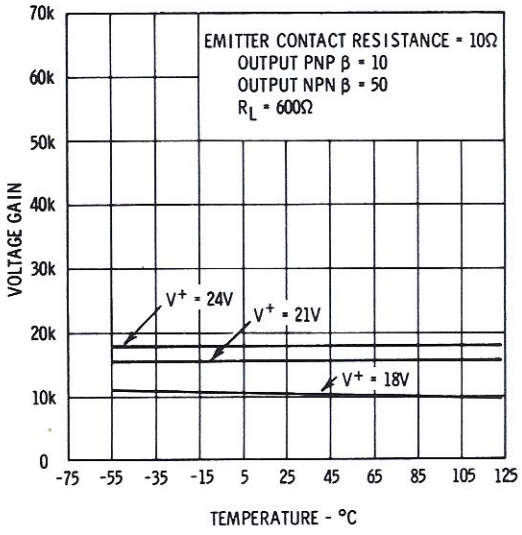


Fig. 7. Voltage gain as a function of ambient temperature.

devices was assumed to be 50 and the load resistor, 600 ohms. The plots cover the range from -55 to 125 degrees centigrade for supply voltages of 18, 21, and 24 volts. The typical variation of emitter contact resistance and initial beta is such that Figs. 3 and 4 can be considered nominal. Figures 1 and 4 through 6 are included to show the relative significance of the uncontrollable parameters on the open loop voltage gain of the device. From the plot set of Fig. 3, the voltage gain variation is seen to be approximately ±1.5 db with temperature and +1.5, -3 decibel with supply voltage (referred to 25°C and +21 V, respectively). Under all operating conditions, the open-loop gain of the μA716, gives a typical closed-loop gain stabiltiy of ±0.25 decibel for temperature and supply voltage variations. Measured voltage gain stability agrees quite well with the predicted value, as shown in Figs. 8 and 9. Figure 8 illustrates the voltage gain variation with temperature while Fig. 9 is a function of supply

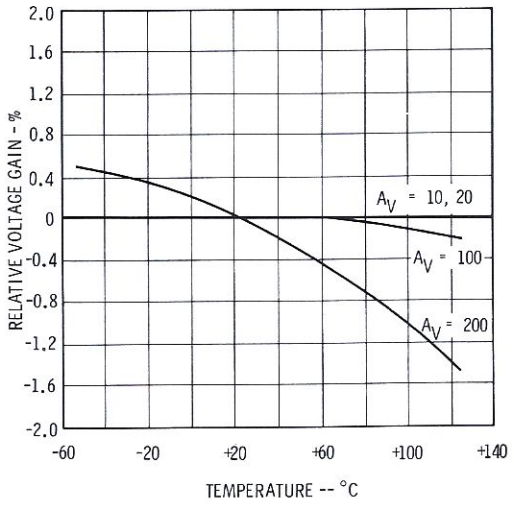


Fig. 8. Relative voltage gain as a function of ambient temperature.

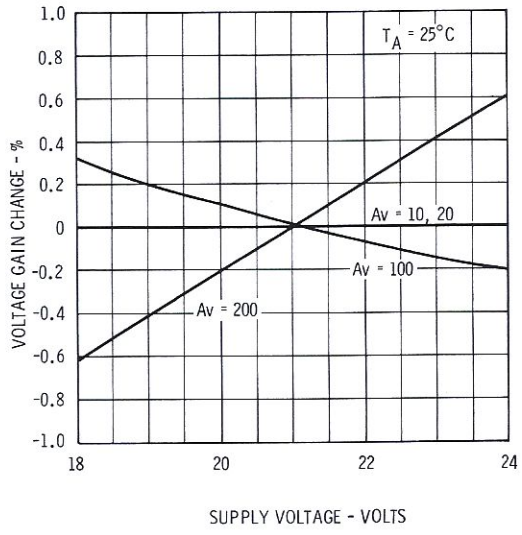


Fig. 9. Relative voltage gain as a function of supply voltage.

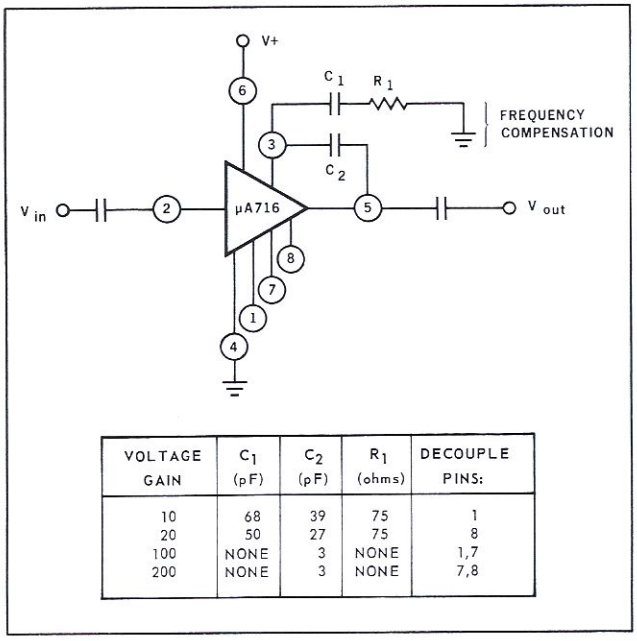
voltage. Variations in closed-loop response for the lowest gain options are generally an order of magnitude better than 200 and 100.

FREQUENCY RESPONSE

The open-loop frequency response of the μA716 is characterized by a three-pole roll-off plus an excess phase term which implies instability for large return differences.[1] Indeed, for the general purpose amplifier, a minimum closed-loop gain of 56 decibels is all that may be obtained without compensation. To stabilize the amplifier for the available gain options, therefore, some external frequency compensation is required. This is provided by a small capacitor connected from the compensation terminal (collector of Q_2) to the output. An additional RC network is necessary for the low gain options (see Fig. 10). The effect of the compensation networks of Fig. 10 on the amplifier response is indicated by Figs. 11 through 15. Figure 11 shows the voltage gain versus frequency characteristic, and Figs. 12 through 15 illustrate the excellent pulse response of the device. From Fig. 11, note how the addition of a small capacitor (3 picofarads) has caused the rolloff of the amplifier to approach the ideal 6 decibels per octave slope. This effect has been referred to as "pole-splitting" by some authors but is better described by the well-known Miller multiplication effect.[2]

1. L. J. Savant, "Control System Design", McGraw-Hill; 1964.

2. L. Strauss, "Wave Generation and Shaping", McGraw-Hill, Section 7-3, pp 214-216.



VOLTAGE GAIN	C_1 (pF)	C_2 (pF)	R_1 (ohms)	DECOUPLE PINS:
10	68	39	75	1
20	50	27	75	8
100	NONE	3	NONE	1,7
200	NONE	3	NONE	7,8

Fig. 10. Connection diagram and component table for available gain options.

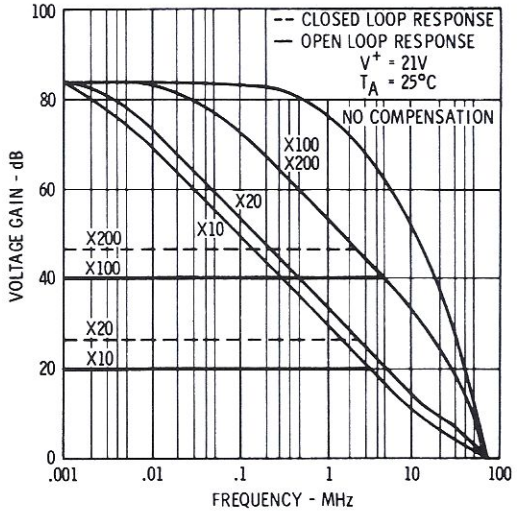


Fig. 11. Open-loop frequency response of amplifier for suggested closed-loop compensation.

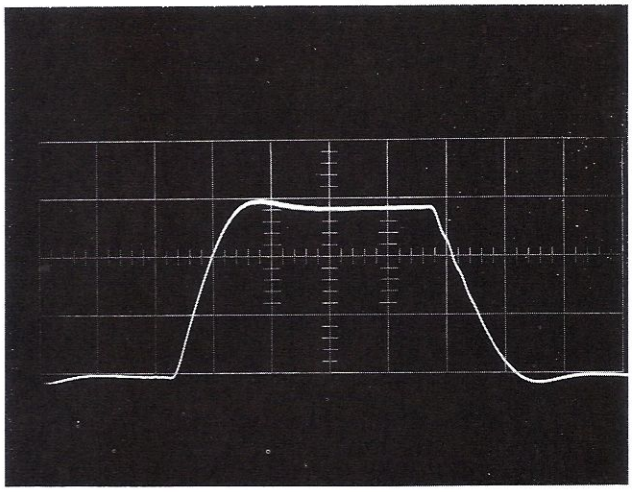


Fig. 12. Pulse response of the μA716 for A_V = 10.

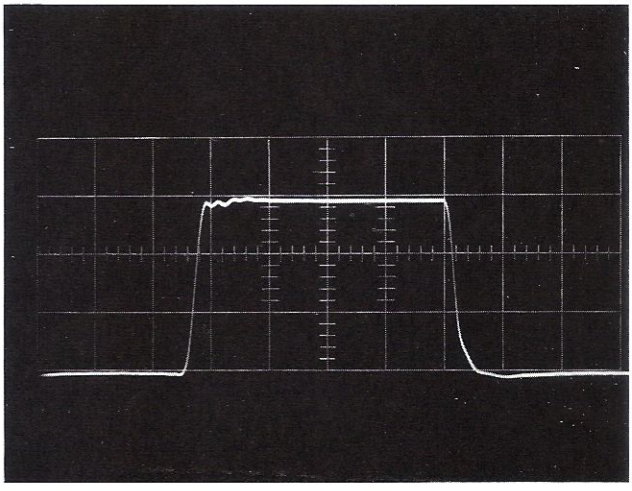


Fig. 13. Pulse response of the μA716 for A_V = 100.

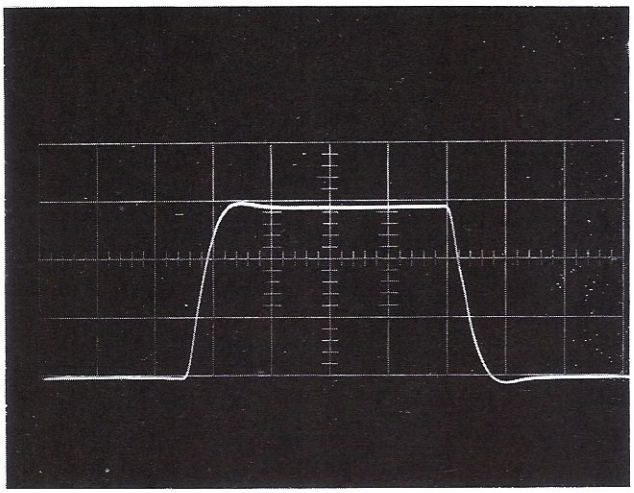


Fig. 14. Pulse response of the μA716 for A_V = 20.

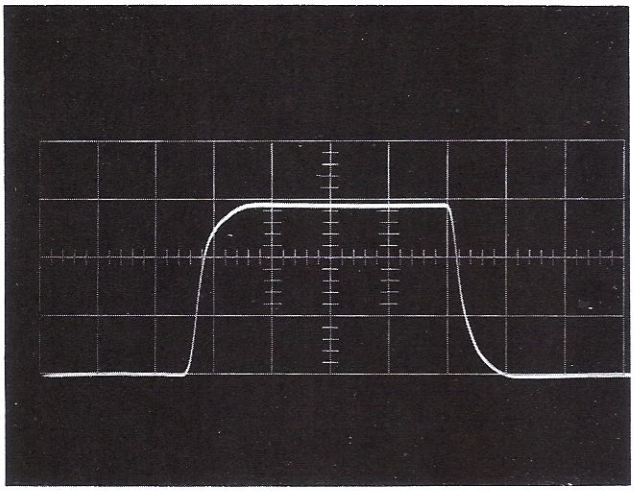


Fig. 15. Pulse response of the μA716 for A_V = 200.

An additional parameter of importance directly related to the bandwidth of the amplifier is the available output power as a function of frequency. Ideally, the maximum available output power of the amplifier should be obtainable over the entire bandwidth of the device. In practice, however, because of stray capacitance and finite current capability at any given node, full power is usually available over only a small fraction of the maximum bandwidth. With operational amplifiers, for example, this limitation is described by a slew rate figure. In the case of an audio amplifier, however, slew rate is not a useful parameter; it is more useful to know the frequency dependence of the output power at a given level of distortion. In Fig. 16, curves of constant harmonic distortion are provided for a 1-percent and 5-percent THD as a function of frequency, parametric in amplifier voltage gain. This shows that maximum output power may be obtained up to approximately 100 kHz.

As mentioned previously, the distortion of the amplifier is present because of the variation in current gain of the output devices with loading. Since large return differences are

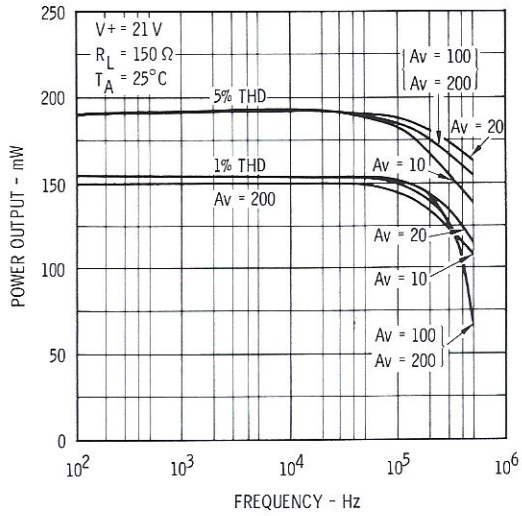


Fig. 16. Power output as a function of frequency.

typical in the device, the nonlinearities in the transfer function of the closed loop amplifier are minimal. A quantitative illustration of the effect of loading on the amplifier is given in Fig. 17 which shows total harmonic distortion as a function of output power parametric in voltage gain. For harmonic distortion less than 0.05 percent, the measurements are essentially equipment limited. The output resistance of the amplifier is a function of the closed-loop gain but is less than 1 ohm for all operating conditions.

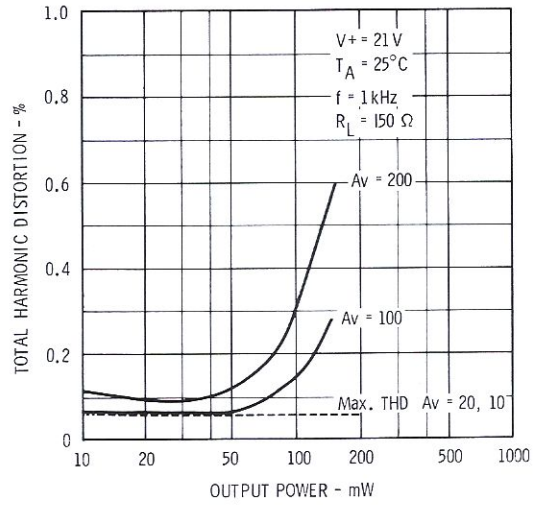


Fig. 17. Total harmonic distortion as a function of output power.

APPLICATIONS

Channel Amplifier

The transmission of voice frequencies between subscribers to a telephone system is generally accomplished by suppressed-carrier, AM modulation. The components of the modulating waveform are limited to frequencies extending

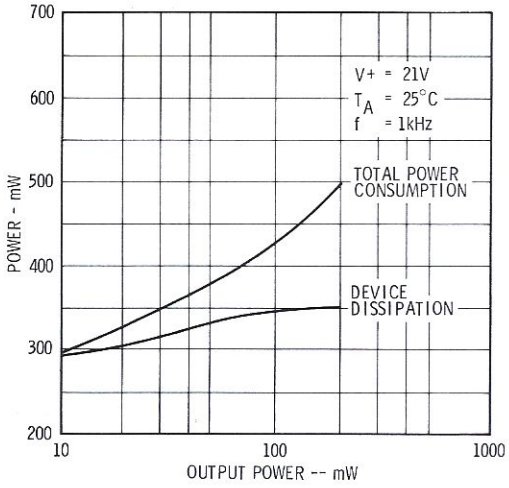


Fig. 18. Total power consumption and internal device dissipation as a function of output power.

from approximately 100 Hz to 3.2 kHz. Upon demodulation at the receiving end of the system, the voice frequencies are amplified, processed, and sent on to the subscriber by a channel amplifier. Electrical specifications for the channel amplifier require low distortion, flat frequency response over the channel bandwidth, 600-ohm input and output impedances to match the transmission line and filters, and an output capability of +13 dBm. In addition, considerable savings can be achieved if dc is eliminated from the output coupling transformer.

The circuit of Fig. 19 shows how the μA716 may be used to satisfy the output impedance, gain, and other requirements of the typical telephone system channel amplifier. The addition of a single external 3-ohm resistor in the feedback network increases the output impedance from 1 ohm to 600 ohms. The desired specifications of a channel amplifier are listed in Table I along with the measured performance of the circuit of Fig. 19.

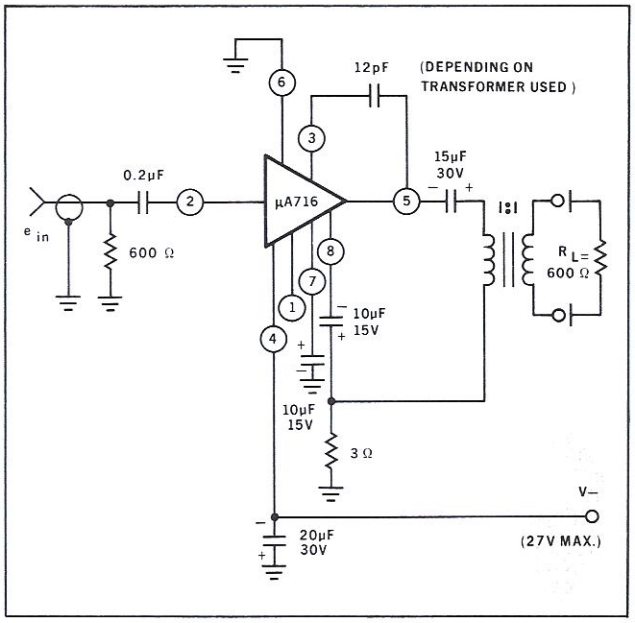


Fig. 19. Channel amplifier connection diagram.

Headset Amplifier

The use of the μA716 as a headset driver may be accomplished very simply as shown in Fig. 20. The potentiometer at the input of the amplifier adjusts the input level while the decoupling capacitor sets the voltage gain of the

TABLE I.

Channel Amplifier Specifications.

PARAMETER		REQUIRED	MEASURED
Bandwidth		100 Hz - 4 kHz	50 Hz - >100kHz
Voltage gain		40 dB	40 dB
Gain stability		±0.3 dB	±0.05 dB
Total harmonic distortion	P_O = +7 dBm	0.5%	0.15%
Input resistance		600Ω ±10%	600Ω ±10%
Output resistance		600Ω ±10%	600Ω ±10%
Output noise power	R_S = 600Ω		
	B_n = 4 kHz		
	A_v = 40 dB	-64 dBm	-75 dBm
Output voltage breakpoint		+13 dBm	+17 dBm
Power supply		-21 V ±1 V	18-24 V
Total power consumption	P_O = 0	–	280 mw
	P_O = 20 mw	< 500 mw	300 mw
Output coupling		ac	ac

Note: (V^- = 20 V, R_L = 600Ω, 0°C< T_A < 70°C.)

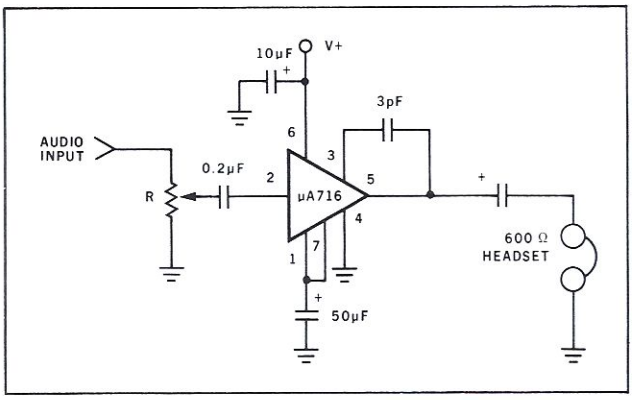


Fig. 20. Headset driver.

amplifier to 100. The 3-pF capacitor from pin 3 to pin 5 of the amplifier provides sufficient frequency compensation to stabilize the device. Use of an additional 3-ohm resistor matches the headset directly to the amplifier, if necessary (Appendix II). If this arrangement is used, however, it is recommended that an additional decoupling capacitor be used at pin 7 (Fig. 21). Such an amplifier may be advantageously used in comunications stations, aircraft or language laboratories.

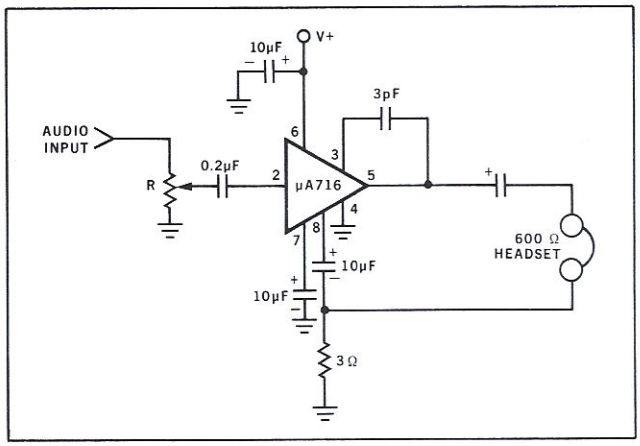


Fig. 21. Driver-connected to match 600-ohm headset.

Medium-Power Amplifier

The μA716C can also be used in dc as well as ac applications as a medium power output stage. Figure 22 shows how it can be used with a μA702C to form an operational amplifier having the low input offset and drift of the μA702C and the low distortion and high output current capability of the μA716C. The open-loop voltage gain of the composite amplifier is 36,000, the closed-loop gain is 200, and the typical input offset voltage is 1.5 millivolts. The amplifier can deliver 15 volts peak-to-peak to a 150-ohm load with less than 0.5 percent distortion from dc to 10 kHz.

Frequency compensation is provided by C_1 -R_4 for the μA702C and by C_2 -C_3 -R_6 for the μA716C.

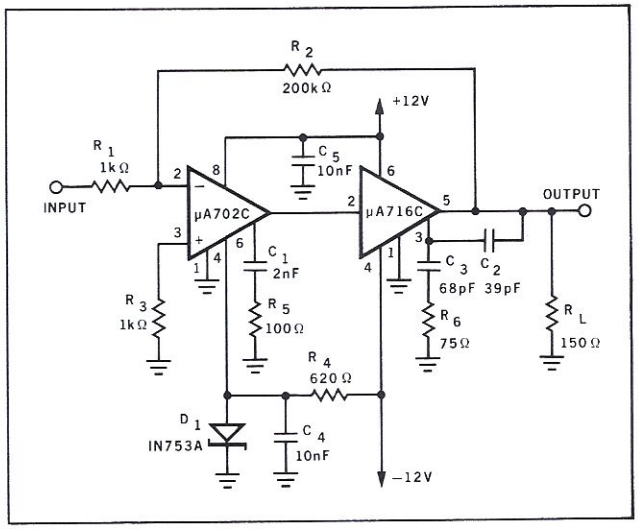


Fig. 22. Medium-power amplifier.

APPENDIX I.

Communications Amplifier Input Stage

From Fig. 23 and loop equations, the following relationships may be obtained:

$$e_o = i_o R_6 = -\alpha_2 i_{e_2} R_6 \tag{1}$$

$$\begin{aligned} e_{in} = &\left[(1-\alpha_1)\ r'_{b_1} + r_{e_1}\right] i_{e_1} \\ &+ \left[(1-\alpha_2)\ r'_{b_2} + r_{e_2}\right] i_{e_2} \\ &- \left[(1-\alpha_3)\ r'_{b_3} + r_{e_3}\right] i_{e_3} \\ &- \left[(1-\alpha_4)\ r'_{b_4} + r_{e_4}\right] i_{e_4} \end{aligned} \tag{2}$$

and

$$\begin{aligned} &\left[(1-\alpha_2)\ r'_{b_2} + r_{e_2}\right] i_{e_2} \\ &= \left[i_{e_1} - (1-\alpha_2)\ i_{e_2}\right] R_4 \end{aligned} \tag{3}$$

where

i_o = output current

r'_b = base spreading resistance

α = common base short circuit current gain

i_e = small signal emitter current

r_e = small signal emitter resistance

$= \dfrac{kT}{qIe}$

$\approx 25/Ie$ ohms

I_e = emitter current mA (dc)

Assuming the devices are matched and the circuit is balanced, the following approximations may be made:

$$\alpha_1 = \alpha_2 = \alpha_3 = \alpha_4 = \alpha$$

$$r'_{b_1} = r'_{b_2} = r'_{b_3} = r'_{b_4} = r'_b$$

$$r_{e_1} = r_{e_4}$$

$$r_{e_2} = r_{e_3}$$

$R_4 = R_5$ (R_4 and R_5 are much greater than r'_b or r_e)

$$i_{e_1} = -i_{e_4}$$

$$i_{e_2} = -i_{e_3}$$

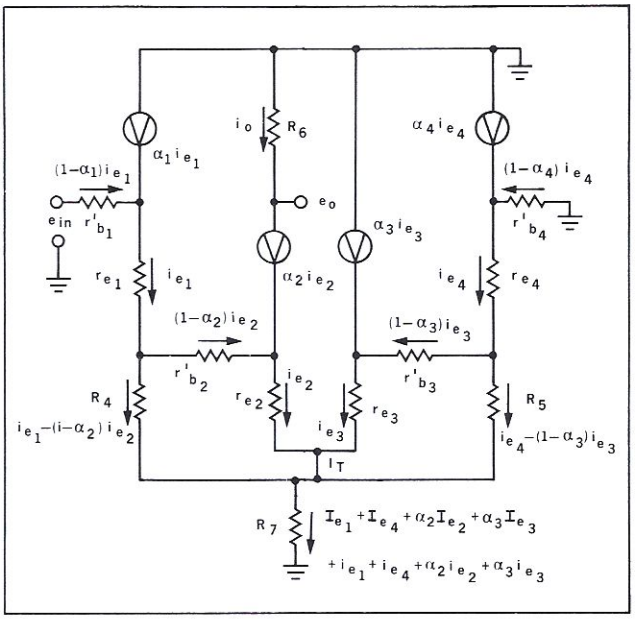


Fig. 23. Small signal equivalent circuit of input stage.

Substitution of the approximations into Equations 2 and 3 yields:

$$\begin{aligned} e_{in} = 2 \Big\{ &\left[(1-\alpha)\ r'_b + r_{e_1}\right] i_{e_1} \\ &+ \left[(1-\alpha)\ r'_b + r_{e_2}\right] i_{e_2} \Big\} \end{aligned}$$

$$\frac{i_{e_1}}{i_{e_2}} = \left[\frac{(1-\alpha)\ R_4 + r_{e_2}}{R_4}\right]$$

$$\cong (1-\alpha) \tag{4}$$

for $(1-\alpha)\ R_4 \gg r_{e_2}$ (5)

$$\frac{e_o}{e_{in}} = -\frac{1}{2} \frac{\alpha R_6}{\left[(1-\alpha)\ r'_b + r_{e_1}\right] \frac{i_{e_1}}{i_{e_2}} + \left[(1-\alpha)\ r'_b + r_{e_2}\right]}$$

$$= -\frac{1}{2} \frac{\alpha R_6}{(1-\alpha)\ r_{e_1} + r_{e_2}} \tag{6}$$

for $(1-\alpha)\ r'_b \ll r_e$

Recalling that

$$r_{e_n} = \frac{kT}{qI_{en}}$$

and that, for the Darlington connection,

$$I_{e_1} = (1 - \alpha)\ I_{e_2}$$

$$r_{e_1} = \frac{r_{e_2}}{1 - \alpha}$$

Equation 6 becomes

$$\frac{e_o}{e_{in}} = \frac{1}{2}\,\frac{\alpha R_6}{2r_{e_2}} \qquad (7)$$

It is evident from Equation 7 that the voltage gain of the Darlington connection is one-half that of the transistor pair. Intuitively, one can see that since the source (the output impedance of Q_1) is equal to the load (the input impedance of Q_2) the voltage developed from base to emitter of Q_2 is less than the source voltage by the factor

$$\frac{\beta r_{e_2}}{\beta r_{e_2} + r_{e_1}}$$

so that r_{e_1} must be made much less than βr_{e_2} in order to realize a gain from the quasi-Darlington connection, approaching that of the transistor pair.

Rather than use resistors R_4 and R_5, current sources may be utilized to bias Q_1 and Q_4 at a large enough current to make $r_{e_1} \ll \beta r_{e_2}$. However, the space required on the integrated structure does not justify their use, particularly since common-mode rejection is not a parameter of interest.

APPENDIX II.

Communications Amplifier Output Impedance

The output impedance of a high gain linear amplifier with negative feedback is well known[1] and is given by the expression

$$Z_o = Z_{po}/(1 - a_o f_o) \qquad (1)$$

or

$$Z_o = A_{po}\,(1 - a_o f_o) \qquad (2)$$

for shunt or series feedback, respectively, where

Z_o = output impedance of amplifier with feedback

Z_{po} = output impedance of amplifier without feedback

a_o = open-loop gain of amplifier

f_o = return ratio of feedback network

For systems requiring an amplifier with a relatively low or high output impedance, feedback methods are available which are easy to reduce to practice. Since the absolute value of the output impedance is typically of secondary importance (i.e., one usually requires something less than 50 ohms or greater than a few hundred ohms), other system requirements will dictate the particular amplifier/feedback configuration used.

Feedback amplifiers are occasionally required to have an output impedance which provides minimum return loss from the load. Such an amplifier, for example, may be required to have an output impedance of 600 ohms for matching to a telephone transmission system. For this class of amplifiers, a combination of shunt and series feedback is, at least intuitively necessary.

The definition of impedance matching that best suits our purpose is that impedance which, when connected to the source, causes the unloaded voltage to be reduced by a factor of two. To see the relationship which exists between a basic feedback amplifier, its closed-loop voltage gain, input voltage, loaded and unloaded output voltage, refer to Fig. 24 and basic power transfer equations ("loaded" and "unloaded" are used here rather than "open-circuit" since the amplifier we finally consider will never, in fact, be "open-circuited". It will be operated without an external load, however.)

From Fig. 24,

$$P_{L(max)} = \left.\frac{e_{oc}^2}{4R_o}\right|_{R_L = R_o} \qquad (3)$$

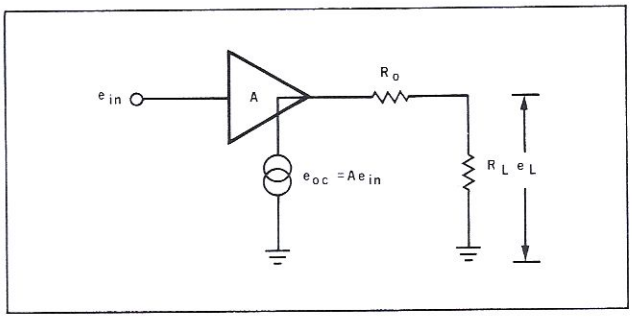


Fig. 24. Basic feedback amplifier with load.

1. Bode, "Network Analysis and Feedback Amplifier Design", D. Van Nostrand Co., Inc.; Fourteenth Printing.

or

$$P_{L(max)} = \left(\frac{e_{oc}}{2}\right)^2 \frac{1}{R_L} = \frac{e_L^2}{R_L} \tag{4}$$

so that

$$e_L = \frac{e_{co}}{2} = \frac{A}{2} e_{in} \tag{5}$$

when matching has been accomplished. Clearly, then, the definition above is correct. The result of Equation 5 may appear trivial, but consider the view of the feedback system under these conditions. If the input voltage to the amplifier remains constant for all load conditions, then the loaded output voltage of the amplifier will be equal to one-half the unloaded voltage if the addition of the load causes the closed-loop gain, A, to be reduced by a factor of two. In other words, if the basic amplifier has an output impedance approaching zero, by proper choice of feedback conditions it can be made to have any finite output impedance.

To illustrate the technique implied by the preceding discussion, an amplifier will be shown which has a loaded voltage gain of 100, an unloaded gain of 200, and an output impedance which matches the load within a fraction of a percent. Perhaps the easiest way to show that matching has been accomplished is to take an amplifier with a known output impedance and the same unloaded voltage gain as above and relate that output impedance to the parameters of the amplifier in question. (The output impedance may also be found from the partial derivative of the output power with respect to R_L, but the mathematics involved provides no clear picture of what the amplifier is doing.)

To begin, consider Fig. 25, which shows the amplifier of interest. If the assumption that the effects of the input and output impedances (R_{in} and R_g respectively) of the amplifier are negligible, then the following may be obtained:

$$\varepsilon = e_{in} - (R_2 + R_3')\, e_o/(R_1 + R_2 + R_3') - R_3''\, e_o/(R_3'' + R_L) \tag{6}$$

$$a_o \varepsilon = e_o \tag{7}$$

$$\frac{e_o}{e_{in}} = \frac{a_o}{1 - a_o (f_A + f_L)} \tag{8}$$

$$\frac{e_L}{e_{in}} = \frac{R_L}{R_L + R_3} \left[\frac{a_o}{1 - a_o (f_A + f_L)}\right] \tag{9}$$

where

a_o = open-loop voltage gain of the amplifier

f_A = amplifier return ratio

$$= \frac{-(R_2 + R_3)}{R_1 + R_2 + R_3}$$

f_L = load return ratio

$$= \frac{-R_3}{R_3 + R_L}$$

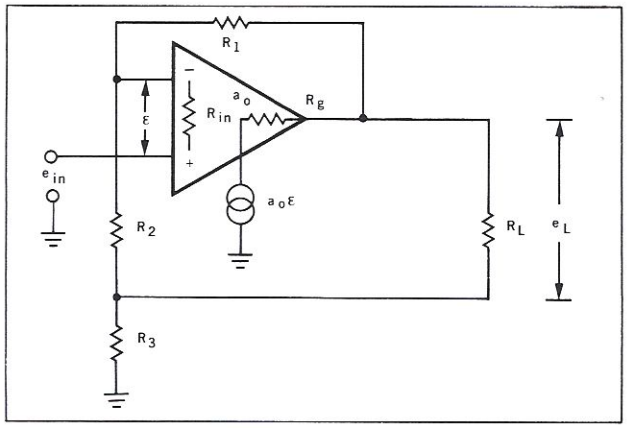


Fig. 25. Amplifier of interest, with feedback.

and

R_3' = R_3 in parallel with R_L

$\approx R_3$

R_3'' = R_3 in parallel with R_2

$\approx R_3$

Similar relationships may be derived from Fig. 26, the amplifier with defined voltage gain and output impedance. They are:

$$\varepsilon = e_{in} - R_{f_2}\, e_o/(R_{f_1} + R_{f_2})$$

$$a_o \varepsilon = e_o \tag{10}$$

$$\frac{e_o}{e_{in}} = \frac{a_o}{1 - a_o f_A'} \tag{11}$$

$$\frac{e_L}{e_{in}} = \frac{R_L}{R_L + R_O} \left[\frac{a_o}{1 - a_o f_A'}\right] \tag{12}$$

where

f_A' = return ratio of the amplifier

$$= \frac{R_{f_2}}{R_{f_1} + R_{f_2}}$$

= f_A (Fig. 25)

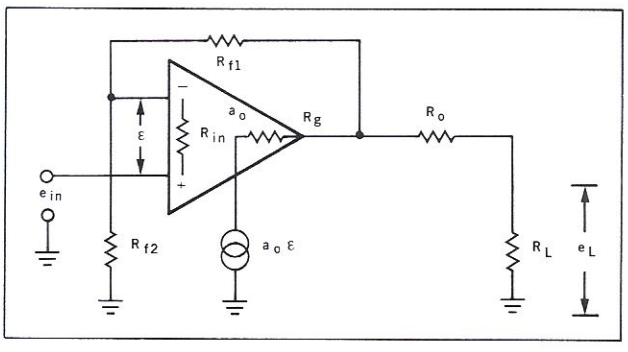


Fig. 26. Amplifier with defined gain and output impedance, R_0.

Equating the input-to-load voltage gain expressions derived from Figs. 25 and 26, solving for R_0 in terms of f_A and f_L, and noting that

$$R_3 + R_L = \frac{R_3}{f_L} \tag{13}$$

the output impedance, R_0, becomes

$$\begin{aligned} R_0 &= R_3 \left[1 - \frac{a_o f_L}{1 - a_o f_A} \right] \\ &\quad - R_L \frac{a_o f_L}{1 - a_o f_A} \\ &= R_3 + (R_3 + R_L) \frac{a_o |f_L|}{1 + a_o |f_A|} \\ &= R_3 \left\{ 1 + \frac{1}{f_A} \left[\frac{a_o f_A}{1 + a_o f_A} \right] \right\} \\ &= R_3 \left[1 + \frac{1}{|f_A|} \right] \\ &= (1 + A)\ R_3 \end{aligned} \tag{14}$$

where

$|f_A|$ = absolute value of return ratio

A = closed-loop gain of the amplifier (unloaded).

Once the closed-loop gain has been specified and the required output impedance found, the application of Equation 14 will define the value of R_3 required, subject to the condition that the amplifier return difference, f_A, is not significantly changed by the addition of R_3. The output impedance in fact is proportional to the closed-loop gain, but since in the μA716, R_2 varies from 365 to 1400 ohms for the different gain options and the closed-loop gain is set by predictable resistor ratios, a wide range of adjustment of output impedance is available.

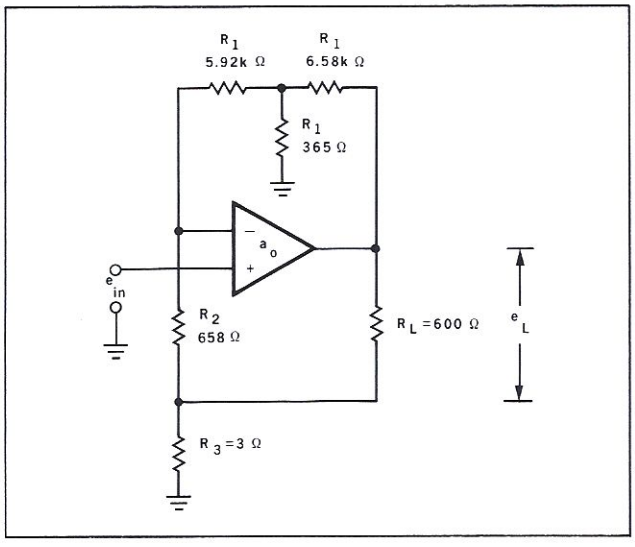


Fig. 27. Communications amplifier connected for A_V = 100 and R_0 = 600Ω.

The actual internal feedback network of the μA716 uses a two-step divider in order to reduce the size of the resistors required. Figure 27 shows the basic circuit connected for a gain of 100 and an output impdeance of 600 ohms.

Feedback Network

The conditions that must be satisfied by the feedback network are: (1) the bias current for Q_4 must be obtained from a 12.5 Kilohms source in order to make the offset voltage as low as possible, and (2) the network must provide unity dc transfer but must be capable of being decoupled to provide return ratios of approximately 200:1, 100:1, 20:1, and 10:1. Additionally, the number of connections is limited to three so that the integrated amplifier can be assembled in an eight-pin TO-5 header.

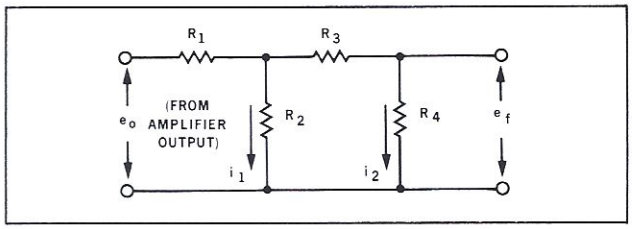


Fig. 28. Equivalent circuit of amplifier feedback network.

Assuming that the effect of input and output impedances of the amplifier are negligible, we can write, from Figure 28:

$$i_1 (R_1 + R_2) - i_2 R_2 = e_o \tag{15}$$

$$-i_1 R_2 + i_2 (R_2 + R_3 + R_4) = 0 \tag{16}$$

Using determinants and solving for i_2

$$i_2 = \frac{\begin{vmatrix} R_1 + R_2 & e_o \\ -R_2 & o \end{vmatrix}}{\Delta} \tag{17}$$

where

$$\Delta = \begin{vmatrix} (R_1 + R_2) & -R_2 \\ -R_2 & (R_2 + R_3 + R_4) \end{vmatrix}$$

$$= (R_1 + R_2)(R_2 + R_3 + R_4) - R_2^2 \tag{18}$$

$$i_2 = e_o \frac{R_2}{\Delta} \tag{19}$$

$$\begin{aligned} i_2 R_4 &= e_f \\ &= e_o \frac{R_2 R_4}{\Delta} \end{aligned} \tag{20}$$

so that

$$\begin{aligned} -\frac{e_f}{e_o} &= -\frac{R_2 R_4}{\Delta} \\ &= f_A \\ &= \text{return ratio of the amplifier.} \end{aligned} \tag{21}$$